THE GW4BWE VERSAFILTER

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THE VERSAFILTER (versatile active filter) is the product of a quest for an audio filter design that would reconcile the requirements of high performance and operational flexibility with the potentially conflicting desire to present a straightforward home construction project that does not feature a control panel resembling the flight deck of Concorde! The result, arrived at after much head scratching, utilizes separate, cascaded lowpass and highpass filters that are independently tunable over the range 300Hz–3kHz, and which facilitate the synthesis of an infinite variety of bandpass responses—giving the operator complete control over both bandwidth and centre frequency. These features enable the optimum filter characteristic for nearly every conceivable mode, including ssb, sstv, cw, rtty and data, to be readily obtained. Despite such flexibility, the system requires only three front-panel controls.

How it works

Readers will be familiar with the concept of a bandpass response characteristic which, in the simplest realization, is obtained using a single-tuned circuit comprising just one inductor and one capacitor. The acceptor circuit thus formed will allow a signal at the circuit's resonant frequency to pass relatively unattenuated, but signals having either lower or higher frequencies are reduced in amplitude by a factor dependent upon the tuned circuit Q and the degree of frequency spacing; ie, how much lower or higher than the resonant frequency the off-tune signals are.

Within the i.f amplifier of a traditional superhet receiver, a number of tuned circuits are cascaded in order to increase the attenuation of adjacant, unwanted signals—hence the use of "shape factor" as a means of expressing the rejection ratio thereby obtained. Contemporary receivers will typically employ either quartz crystals or ceramic resonators to act as high-Q substitutes for LC tuned circuits, but the principle is exactly the same.

It is also possible to reproduce the tuned circuit bandpass response at audio frequencies using an active filter. Active filters allow us to dispense with bulky inductors by employing a clever circuit arrangement, known as the gyrator, which enables a humble capacitor to assume the role of an inductor. Active cw filters that are designed to be interposed between the receiver's audio output socket and an external loudspeaker or headphones have proved very popular. Cascading a number of simple bandpass stages (each being the direct equivalent of a low-Q tuned circuit) enables a respectable shape factor to be obtained, and by slightly off-setting the resonant frequencies of each stage we can arrive at a wider, flat-topped passband response [1]. A further design option involves the provision of a system for tuning a number of cascaded bandpass stages in unison. This allows the centre frequency of the entire filter to be shifted over a considerable range [2].

Although the construction of active filters intended specifically to be used for the enhancement of cw reception is both a practical and cost-effective pursuit, many operators are likely to demand a filter system which boasts a much wider range of applications. For instance, let's assume that aside from cw reception we also wish to improve the intelligibility of ssb signals during operation in crowded phone segments of popular hf bands. A high-performance audio filter featuring a much wider bandpass response, perhaps extending over the range 300–2,700Hz, would clearly be of value in rejecting adjacent channel interference. Of course, the degree of improvement obtainable depends, to some extent, on how good a shape factor the receiver's i.f filter possesses. Nevertheless, even if a high-quality eight-pole, 2·4kHz crystal filter is fitted in the station receiver/transceiver, the level of low-frequency "rumbles" and high-pitched "chatter" will still be significantly reduced by a decent audio filter.

We now have to ask whether it is possible to configure an active bandpass filter which can be varied in bandwidth from less than 100Hz at −6dB for cw reception under difficult conditions, to around 2·5kHz for phone operation. Also, there will be a requirement for a variety of intermediate bandwidths and centre frequencies so that the same filter may also cope with rtty and data modes. This is quite a tall order, especially if we further demand that a good shape factor is preserved at all settings.

The active cw filters discussed above can provide some flexibility, but the increase in bandwidth obtainable by, for instance, lowering the Q of each stage is, in the context of our stated requirement, only moderate. Furthermore, if Q is reduced to a value lower than about two, a large number of cascaded stages are required in order to achieve a respectable shape factor. There is, however, an alternative approach. This involves the synthesis of a bandpass response using separate lowpass and highpass filters that are simply cascaded.

As an example, let's imagine that the highpass filter has a cut-off frequency of 500Hz and the lowpass filter a cut-off frequency of 1kHz. Frequencies below 500Hz will be attenuated by the highpass filter and, conversely, frequencies above 1kHz are attenuated by the lowpass section. However, signals at frequencies falling between 500Hz and 1kHz will travel through the filters unaffected. We have therefore "synthesized" a bandpass filter of 500Hz bandwidth (1,000 − 500Hz), and provided that the lp and hp sections are correctly designed, a commendably flat passband response is assured.

It is perfectly feasible to make both filters independently tunable over the range 300Hz–3kHz, and as there is no requirement for a variation in the Q of either filter, shape factor is never unduly compromised. Needless to say, the variety of bandwidths and centre frequencies obtainable is infinite: if the hp section is set at 700Hz and the lp section at 800Hz we obtain our 100Hz cw filter, whereas hp = 300Hz and lp = 2·7kHz produces an effective ssb filter.

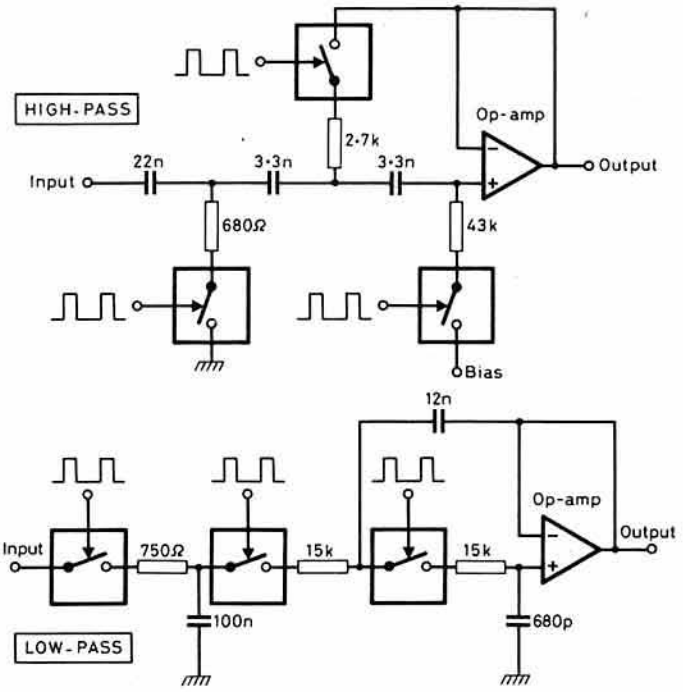

Fig 1. The tunable third-order active filters which form the heart of the versafilter

Design principles

The lowpass and highpass filter circuits which constitute the building blocks for the versafilter are shown in Fig 1. These are third-order types and therefore offer an ultimate attenuation of 18dB/octave. Readers may recognize a similarity between the lowpass circuit featured in Fig 1 and the filter stages employed in a direct-conversion receiver module that I have previously described [3].

Tuning is achieved by making the three resistors used in each circuit variable, so that their values may be altered in unison while maintaining the correct ratios; ie, in the highpass circuit, if we wish to lower the cut-off frequency by one octave, the 680Ω resistor is increased to 1·36kΩ, 2·7kΩ becomes 5·4kΩ, and 43kΩ becomes 86kΩ. In order to make this possible, logic-controlled signal gates, four of which are contained within a single

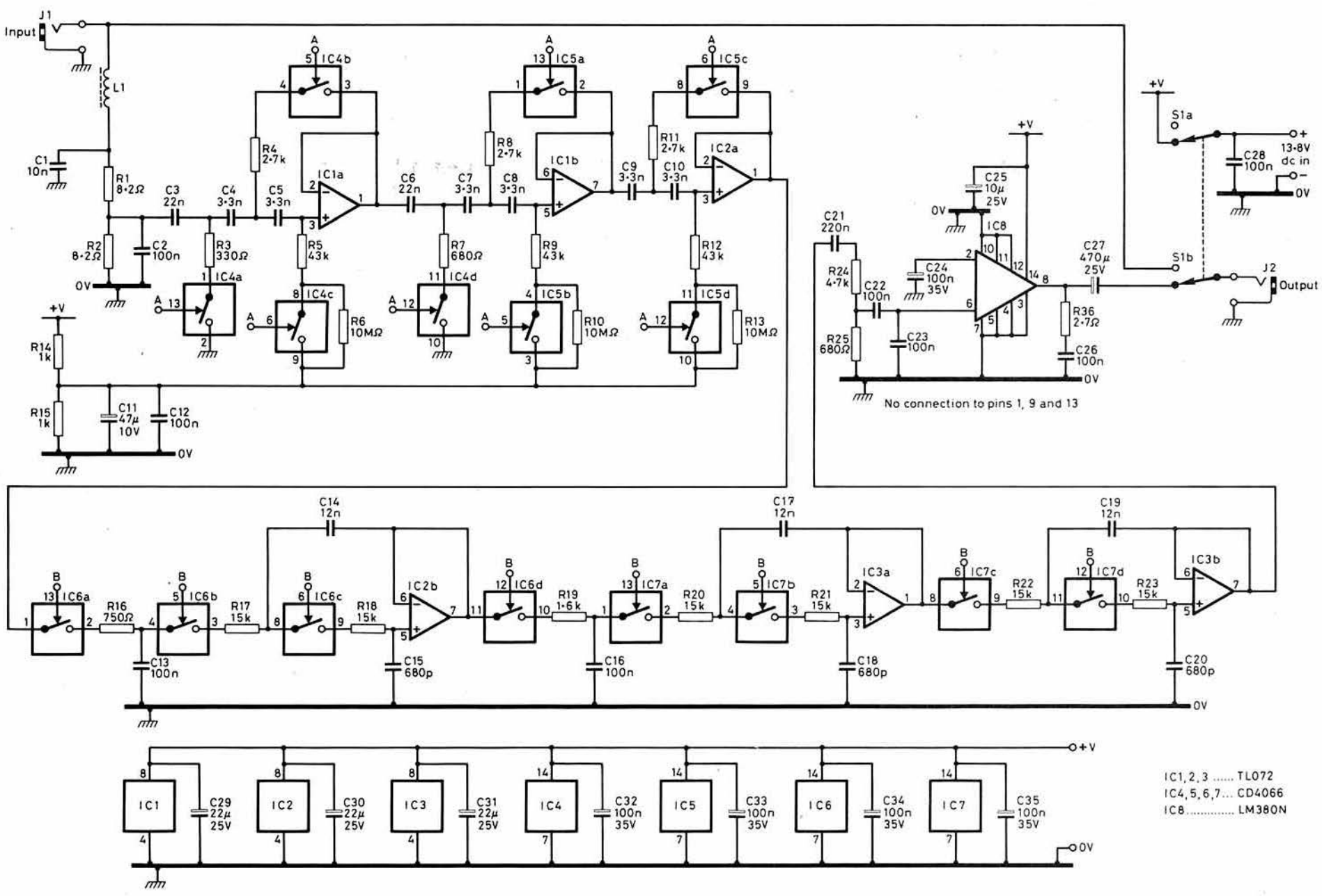

Fig 3. Circuit of the filter system. L1 is fabricated by winding two turns of 24swg enamel copper wire through a twin hole ferrite bead (see components list)

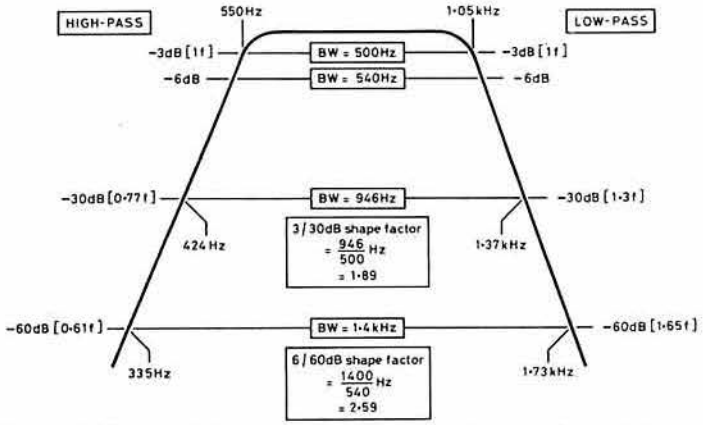

Fig 2. A typical versafilter response curve obtained by setting the −3dB cut-off frequencies of the highpass and lowpass filters at 550Hz and 1·05kHz respectively

CD4066 cmos integrated circuit, are wired in series with each resistor. By successively opening and closing the gates at an ultrasonic frequency, a multiplication of the fixed resistor value is obtained. The degree by which the resistances are multiplied is directly proportional to the mark/space ratio of the switching waveform, so tuning the filters becomes a matter of making this factor variable. The technique just described was also used in my tunable cw filter [2].

Unfortunately, the shape factor of a bandpass response obtained by cascading the two circuits shown in Fig 1 is not nearly good enough. It is practicable, however, to employ a number of third-order sections working in unison. The versafilter therefore uses the equivalent of three lowpass, and also three highpass sections so as to achieve much better performance.

The low and highpass filters must, of course, be made independently tunable over the range 300Hz–3kHz, and this is where the design begins to look rather unwieldly. As we have seen, each third-order section utilizes three "variable resistors", so three such sections would normally require the provision of no less than nine. This is annoying, because a single CD4066 contains four signal gates, and it is obviously more convenient to arrive at a circuit arrangement employing just eight variable resistors; ie, so that only two CD4066s are necessary in the lowpass and highpass filters respectively, making four in total.

Fortunately, a very minor modification allows us to dispense with two of the variable resistors. Each third-order section consists of a passive first-order network followed by an active second-order network. In the lowpass filter (Fig 1) the 750Ω resistor and 100nF capacitor form the first-order network, whereas in the highpass circuit the component positions are interchanged and the values become 680Ω and 22nF. It is possible, by appropriate scaling of values, to modify the cut-off frequency of a first-order network so that it may substitute for two separate first-order networks in a given filter design. This technique is exploited in the versafilter to reduce the total number of first-order networks from six to four. The modified lowpass network consists of 1·6kΩ and 100nF, and the highpass uses 330Ω and 22nF. Note that scaling of the cut-off frequency only involves changing the resistor value in each case.

In contrast to higher-order and/or elliptical arrangements, the uncritical, low-Q third-order design lends itself to home construction using "off-the-shelf" five-per-cent tolerance capacitors and resistors. A very respectable performance is virtually guaranteed, as illustrated by Fig 2 which shows a typical versafilter response curve.

Circuitry and construction

The circuit of the versafilter appears in Figs 3 and 4. Fig 3 shows the complete signal chain which is arranged as follows: The rig's external loudspeaker socket is coupled to the input jack, J1, and signals are routed via L1, C1 (an rf filter) to a 6dB attenuator formed by R1 and 2. The tunable highpass filter follows, and this comprises IC1, IC2(a), IC4 and 5 plus associated resistors and capacitors (R3–R13, C3–C10). The potential divider, R14 and 15, develops a half-rail bias supply which is decoupled by C11 and 12. The lowpass filter is driven from the output of IC2(a), and as the lp sections are essentially direct-coupled, no further bias feeds need be provided.

IC8, an LM380N audio power amplifier, provides a loudspeaker output via C27, S1b. S1a is the ON/OFF switch and therefore has its pole connected directly to the 13·8V dc power input socket. S1b is the by-pass switch and, because it is ganged with S1a, switching the filter off will also simultaneously couple the versafilters input straight through to the output jack, J2. IC8 contributes a voltage gain of approximately 30dB, so the potential divider comprising R24 and 25 is incorporated to provide a degree of attenuation at the input of IC8. The versafilter's net gain is close to unity, so provision of a volume control was considered unnecessary.

The tuning circuitry is shown separately in Fig 4. IC9 operates as a 32kHz clock oscillator with initial frequency setting facilitated by R29. The square-wave output from IC9 is fed to the inputs of IC10, a dual monostable. IC10 has two independent outputs, labelled A and B, which drive the control pins of the CD4066s used in the highpass (A) and lowpass (B) filters. For clarification, Fig 5 provides an overview of the versafilter system.

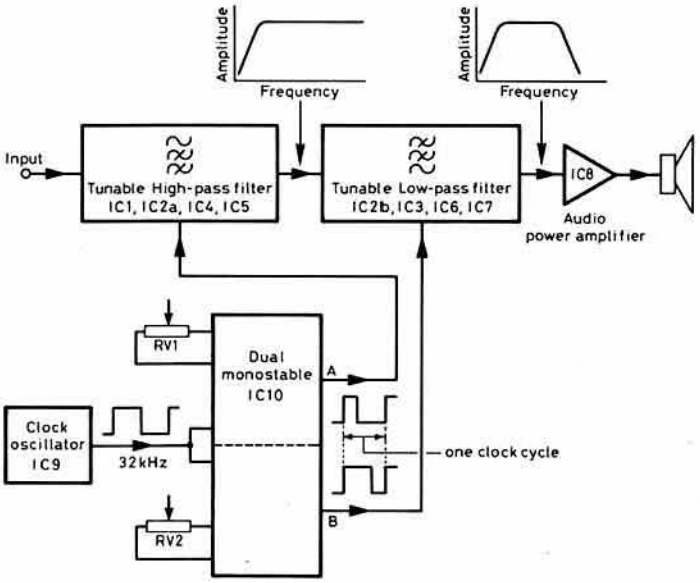

Fig 5. Versafilter block diagram

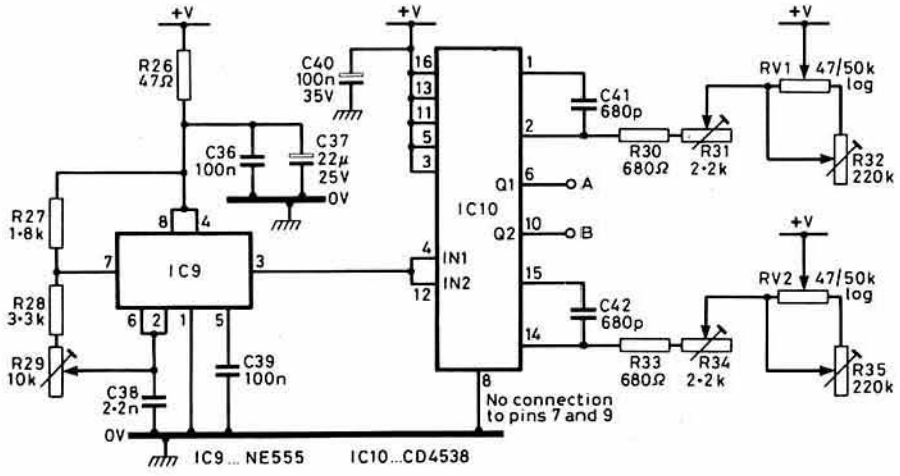

Fig 4. Clock oscillator and dual monostable tuning circuit

Components list

C1	10nF ceramic
C2, 12, 26 28, 36, 39	100nF ceramic
C3, 6	22nF polyester
C4, 5, 7, 8, 9, 10	3·3nF polyester
C11	47μF electrolytic 10V
C13, 16, 22 23	100nF polyester
C14, 17, 19	12nF polyester
C15, 18, 20, 41, 42	680pF polystyrene
C21	220nF polyester
C24, 32, 33, 34, 35, 40	100nF tantalum electrolytic 35V
C25	10 1μF tantalum electrolytic 25V
C27	470μF electrolytic 25V
C29, 30, 31, 37	22μF electrolytic 25V
C38	2·2nF polyester

The polyester capacitors are Siemens B32560, and the polystyrene are Siemens B31310 or similar (5% tolerance).

R1, 2	8·2Ω
R3	330Ω
R4, 8, 11	2·7kΩ
R5, 9, 12	43kΩ
R6, 10, 13	10MΩ
R7, 25, 30, 33	680Ω
R14, 15	1kΩ
R16	750Ω
R17, 18, 20, 21, 22, 23	15kΩ
R19	1·6kΩ
R24	4·7kΩ
R26	47Ω
R27	1·8kΩ
R28	3·3kΩ
R29	10kΩ preset type PR10, 10mm dia
R31, 34	2·2 kΩ preset type PR10, 10mm dia
R32, 35	220kΩ preset type PR10, 10mm dia
R36	2·7Ω

All fixed resistors are 0·33W, 5% tolerance carbon or metal film types (eg, CR25 or UPM033)

RV1, 2	47 or 50kΩ linear slider potentiometers. Cirkit stock code 48-90914
IC1, 2, 3	TL072
IC4, 5, 6, 7	CD4066
IC8	LM380N
IC9	NE555
IC10	CD4538

L1 Siemens twin-hole ferrite bead B62152 A0007 × 030 (See Fig 3 caption for winding details)
S1 Miniature toggle dpdt. Electrovalue S7201 or Cirkit 53-00201/
Case Radio Shack (Archer) aluminium enclosure size 49 by 210 by 156mm, cat no 270-272 available from Tandy.

Miscellaneous
PCB (see text). Sockets for ics: four 8-pin, four 14-pin, one 16-pin dil, input, output and power. Slider knobs for RV1, 2. Mounting/fixing sundries.

Front-panel control of the filter cut-off frequencies is achieved by the use of two linear slider potentiometers, RV1 (highpass) and RV2 (lowpass). But why slider, rather than the more popular rotary type? A glance at Fig 6, which illustrates the recommended front-panel layout, should answer this question. If RV1 and 2 are mounted as shown, the response shape that results from a particular setting is easily visualized. The preset pots, R31, 32, 34 and 35, facilitate calibration of the tuning controls (more about this later). IC8 is soldered in place so as to ensure maximum heat transfer from pins 3–5 and 10–12. The other ics should be mounted in dil sockets of proven quality; eg, Texas low-profile.

Fig 7 gives a rear view of the front panel. The prototype versafilter employs two 50kΩ, 60mm travel, linear slide potentiometers. S1 is a miniature dpdt toggle. The input and output jacks (J1, 2) plus the power socket are mounted on the back panel, and the types employed are left to the constructor's choice. Note that C28 is soldered directly across the power socket.

The capacitor types specified in the components list should be adhered to, and it is certainly not advisable, under any circumstances, to employ ceramics instead of polystyrene or polyester in the filter stages. Finally, the case employed to house the filter must be screened, and aluminium is probably the easiest material to work with (remember that it will be necessary to cut slots in the front panel for the controls of RV1 and 2). Details of a suitable commercial enclosure are given in the components list.

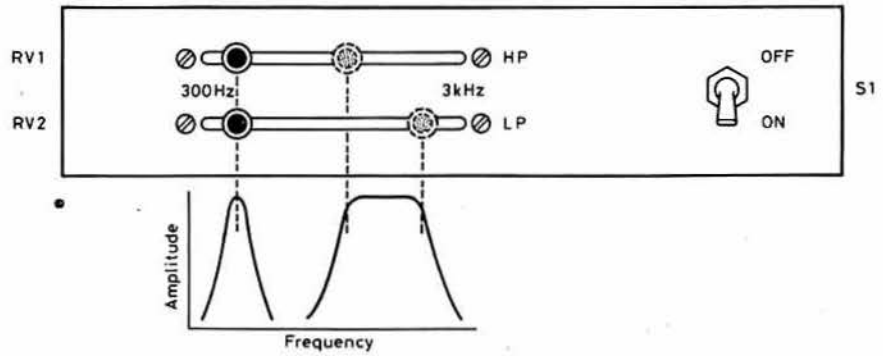

Fig 6. Recommended front-panel layout showing how the slider potentiometers give visual indication of the many response curves obtainable

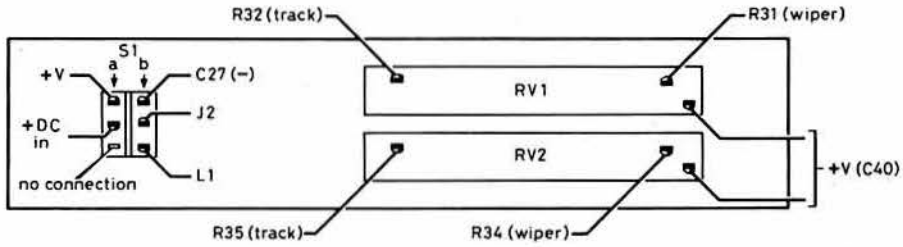

Fig 7. Control panel wiring

Testing and calibration

Constructors who either possess, or have access to, reliable test equipment —including an af signal generator, ac millivoltmeter and oscilloscope— may, of course, employ such items to aid calibration. However, as many amateurs are forced to get by with little more than a multimeter, a simplified commissioning procedure will now be outlined.

(1) A 13·8V psu is coupled to the power socket and, with S1 in the "on" position, the presence of rail voltage at the power pin of each ic is confirmed using a dc voltmeter. A reading of approximately 6·9V, ie, half rail, should be obtained at the junction of R14 and 15.

(2) With R29 set at mid-travel, a dc voltage reading of around half the rail potential at pin 3 of IC9 indicates that the clock oscillator is almost certainly functioning correctly.

(3) A short length of hook-up wire is temporarily connected to pin 3 of IC9. An hf receiver placed alongside the versafilter can now be used to monitor the harmonics generated by the clock oscillator (it may also be necessary to couple a make-shift wire antenna to the receiver). Use the lowest frequency range available and check the spacing between the harmonics by tuning through successive zero beats. The nominal clock frequency required is 32kHz, so R29 must be adjusted to obtain a harmonic spacing close to this. In practice, anything between 30 and 33kHz is perfectly acceptable.

(4) The dual monostable (IC10) is now checked as follows. The presets R31, 32, 34 and 35 are initially set at mid-travel, and a dc voltmeter is connected to pin 6 of IC10. Adjustment of RV1 should cause large fluctuations in the reading obtained—this serves as an indication that the mark/space ratio is indeed being controlled by RV1. The meter is now connected to pin 10 of IC10 and the same test repeated for RV2.

(5) If everything appears satisfactory, the audio output of the station receiver should be coupled to the versafilter input jack, and a loudspeaker connected to the filters' output. Tune the receiver to a busy frequency, ensuring that its af gain control is fairly well advanced. With S1 in the "off" position, the audio will be unaffected. Furthermore, disconnecting the power supply from the filter should make no difference. However, leaving the psu disconnected and flicking S1 to the "on" position will silence the loudspeaker. If this does not occur, check the wiring of J1, J2 and S1.

(6) Reconnect the psu and check that S1 remains in the "on" position. Set RV1 to the extreme left of its travel (300Hz position) and RV2 to the extreme right (3kHz).

The versafilter is now operational, and it should be possible to monitor signals passing through the filter. However, as the presets associated with RV1 and 2 have yet to be trimmed, the intelligibility of phone transmissions may well be affected.

Moving RV2 from right to left will cause a distinct reduction in the level of high frequencies, making the sound very muffled and indistinct.

Returning the control to the extreme right and then moving RV1 from left to right should produce the opposite effect; ie, a progressive and very marked attenuation of low audio frequencies.

(7) It will not be possible to achieve precise calibration of the filter unless an audio signal generator is available. This is not really important, however, and the fact that the frequency range finally obtained may differ from the nominal 300Hz–3kHz is of little consequence. Nevertheless, fairly accurate tracking between RV1 and 2 is desirable so that when both controls are set at the same position (see left-hand setting illustrated in Fig 6) the filter cut-off frequencies coincide, thus producing the expected narrow, bandpass response.

The following procedure enables the versafilter to be aligned subjectively. First, set RV1 to the extreme left and RV2 to the extreme right. Switch S1 "off" so as to by-pass the filter, and tune the receiver to a strong ssb transmission which has good quality audio. Now bring the filter into operation by flicking S1 "on". R31 facilitates adjustment of the highpass minimum cut-off frequency, and this should be set to a point that provides a reasonable low frequency response. Aim for good intelligibility here but expect a noticeable loss of the lower, "bass" sounds, especially if a large loudspeaker is being used. It should be noted that if the combined resistance of R30 and 31 falls below approximately 1kΩ the monostable (IC10) may "lock-up" and produce a constant dc output. This condition can be detected with a voltmeter connected to pin 6, or by listening to the sudden change in audio quality that results if R31 is advanced beyond the critical point; this comment also applies to R34.

Having settled on an adjustment for R31 we can proceed to set R35, which determines the lowpass maximum cut-off frequency. Listen to the high-frequency content of the speech and trim R31 so that there is just a slight attenuation of these upper frequencies. The by-pass facility provided by S1 may, of course, be employed to make instant comparisons between the direct and filtered speech. Finally, R34 (lowpass minimum cut-off frequency) and R32 (highpass maximum cut-off frequency) are set to the same positions as R31 and 35 respectively.

The tracking of RV1 and 2 may be checked by tuning the receiver to a stable cw transmission and peaking the signal in a narrow bandpass response obtained by moving the hp and lp controls in tandem. Using the receiver tuning knob to shift the pitch of the cw enables a check to be made at various frequencies. If the tracking is poor at low frequencies, try adjusting R34. Conversely, use R32 to correct for any irregularities at high frequencies.

NB. The dc voltage readings at pin 3 of IC9 and pins 6 and 10 of IC10 referred to in sections (2) and (4) must be made with an analogue, moving-coil voltmeter/multimeter.

Conclusion

There are so many potential uses for the versafilter that a comprehensive "operators guide" would probably take up as much space as this article.

If you are utilizing a home computer to decode rtty, data or cw (shame on you if it's cw!) the versafilter may be employed to clean-up noisy, QRM-infested signals before they are presented to the interface/terminal unit, thus reducing errors quite significantly. To avoid the effects of agc pumping when there are strong adjacent-channel signals, try reducing the receiver's rf gain—this can sometimes work wonders.

Although the provision of an accurate tuning scale is a difficult task and requires test equipment, constructors may wish to consider the use of markers for important frequencies. For instance, rtty enthusiasts could use a small sticker to indicate the mid-point between 1,275 and 1,445Hz. The correct position is found, initially, by feeding the output of the rtty tone generator into the versafilter and then peaking RV1 and 2 for maximum response. As the filter input impedance is only 16Ω it will probably be necessary to add a series resistor to prevent loading the tone generator excessively.

Although I might be accused of taking liberties if I were to try and describe this design as "simple", it is certainly far less complex than some other fully tunable audio filters. Furthermore, painless construction of this project is made possible by the availability of a high-quality, pre-drilled printed circuit board.

The versafilter pcb, plus accompanying layout diagram, may be obtained from: FAW Electronics, "Four Seasons", Straight Road, Battisford, Stowmarket, Suffolk IP14 2LZ. The price, including VAT and p&p is £12.50.

References

[1] "The G4BWE cw filter". *Rad Com* March 1983, pp226–9.

[2] "The G4BWE tunable cw filter". *Rad Com* September 1984, pp755–8.

[3] "Direct-conversion cw transceivers", *Rad Com* January 1986, pp26–31 and 33. □

A "SPROGGY" HUNT

N D N Belham, G2BKO*

IT ALL STARTED when a neighbour stopped me in Evesham and, looking at my RSGB badge, said, "So you are G2BKO. I listen to you each Monday morning, in fact you wipe out Radio 2 vhf fm on my transistor set." Aghast, I hastened home to listen on a transistor set for myself and found that there was a indeed a signal on 89·3MHz, even when the transmitter was only putting out a third of a watt. So the hunt began.

My 144MHz transceiver is a "black box" which, fortunately, had been supplied with a comprehensive handbook. Fig 1 is a very much simplified block diagram to show how the transmitter frequency is generated. The modulated output from a 10·7MHz crystal oscillator is added to that from a voltage-controlled oscillator which is variable between 133·3 and 135·3MHz. This variation is limited to 10kHz steps by a phase-locked loop.

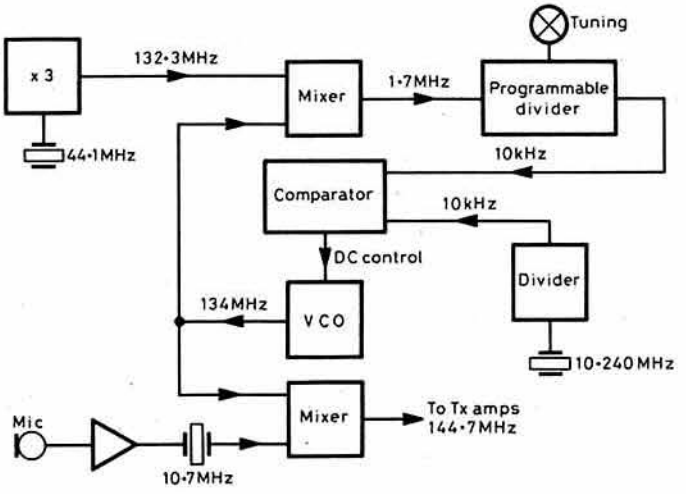

Fig 1. Simplified block diagram of "black box" frequency generating system

Part of the vco output is mixed with the tripled output from a 44·4MHz crystal oscillator to produce a difference frequency between 1 and 3MHz. This difference frequency is reduced by a programmable divider chain set by the tuning control. A comparator stage compares the phase of the two 10kHz signals. The first derived from a 10,240kHz crystal oscillator by a fixed divider chain, and the second from the vco. The comparator supplies the control voltage for the vco.

Thus, when the transmitter is tuned to 144·7MHz, the following frequencies are present in the transmitter: 10kHz, 1·7MHz, 10·24MHz, 10·7MHz (modulated), 44·1MHz, 132·3MHz (3 × 44·1), 134·0MHz (vco) and 144·7MHz, as well as harmonics, summation and difference frequencies!

Could any combination of these produce a modulated signal on 89·3MHz? One could, I suppose, ask a computer, but a little thought and a few simple experiments can limit the possibilities to be considered. Although the i.f of my transistor set was 10·7MHz, direct pick-up from the modulated 10·7 crystal oscillator was ruled out since the unwanted signal was not received all over the band.

The next thing to do was to search for any spurious emission from the transmitter on 89·3MHz, or one having harmonics on that frequency. A Heathkit dip meter was available which consisted of a set of interchangeable coils, a calibrated tuning capacitor and a fet voltmeter (which could be made into an oscillator). A tremendous response was obtained on 144·7MHz when the dip meter was placed near the dummy load (with the transmitter on low power) but none at all on 89·3 or 44·65MHz.

It must be admitted that a dip meter is not as sensitive as a receiver. A more sensitive device was therefore constructed, consisting of a ganged pair of tuned circuits followed by a broadband integrated circuit amplifier. The circuit is shown in Fig 2. The output from the amplifier is rectified by the fet acting as a peak voltmeter. Before any signal is applied, the 5·6kΩ and 4·7kΩ variable resistors are set so that both the 100μA and the voltmeter are reading zero; ie the fet is at cut-off. The peak voltage of the signal applied is then read on the voltmeter by adjusting the 4·7kΩ variable to cut-off

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